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Implementation of Transmit Antenna Weight Adaptation through Stochastic Gradient Feedback

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Abstract - *Feedback assisted transmission antenna weight adaptation has been proposed for Third Generation cellular systems to provide increased capacity on the forward link. This approach provides enhancement concurrently from fading mitigation and beam steering. One such algorithm maximizes the usable power to the mobile by perturbation extracted gradient adaptation of the transmit weight vector. With perfect channel estimation and other idealized assumptions this method has been shown to be capable of providing weights approaching that of the optimal multi-element transmitter. In this paper a method for common perturbation transmission is introduced and compared to dedicated perturbation transmission in a CDMA system. In addition, a subspace weighting approach is introduced to enhance the algorithm performance in correlated fading. Simulations of the algorithm operating in a cdma2000 system incorporate feedback bit errors, latency and channel estimation errors with a dedicated pilot.*

1. Introduction

The use of multiple receive and transmit antennas has been considered for capacity improvement in next generation cellular systems. While space time coding techniques for MIMO (multiple-input multiple-output) systems have recently received great attention, e.g. [1], for reasons of cost and size limitation the number of antennas at the mobile receiver may be limited, and may in fact be limited to one receive antenna. The beneficial use of multiple antennas at the base station in this case is a classical problem. While algorithms for multiple receive antennas at the base station are straight forward and have been thoroughly considered, the use of beam forming with multiple transmit antennas at the base station is complicated in FDD (frequency division duplex) systems by a lack of channel state information at the base station. Blind multi-antenna transmission coding schemes can attain antenna diversity in the MISO (multiple-input single-output) case [2][3], but not the available aperture gain. At the same time, blind multi-antenna transmission

weighting schemes can attain some aperture gain for MISO [4][5], but not diversity gain. This paper develops techniques that provide tracking of transmission weights in order to attain both diversity and aperture gain.

Providing gains from both fading diversity and beam steering in a MISO system requires the use of detailed downlink channel state information. In FDD systems this requires mobile to base feedback. This has led to several proposals for channel state feedback, both in the literature [6][7][8] and in 3rd Generation standardization forums. In [9] a feedback stochastic gradient algorithm was introduced for transmit weight adaptation. The usable power delivered to the mobile is considered to be an inverse cost, which is to be maximized. The proposed system uses weight vector perturbations to extract a coarse estimate of the gradient of this inverse cost, which is then used to adjust the transmit antenna weights. The convergence and dynamic fade tracking of this algorithm with perfect receiver channel estimation and error free instantaneous feedback were considered in [10], where the stochastic gradient feedback technique was shown to outperform diversity space time coding [2] and vector selection feedback [7] in slow fading channels.

In the previous investigations [9][10] the gradient feedback technique required a dedicated perturbation transmission, specific to a single receiver. This requires additional transmission power, which in a DS-CDMA system translates to a loss of capacity through excess interference. In a DS-CDMA system with multiple receivers sharing the same band, transmission of a single common perturbation used by all receivers can minimize the excess interference, similar to the use of a single common pilot in traditional CDMA systems. In the present work, a generalized formulation of the stochastic gradient feedback approach is presented which allows for either dedicated or common perturbation transmission. In addition, this work introduces an approach for subspace weighting of the weight vector update, where a matrix pre-multiplication of the probing or update perturbation vector is performed in order to emphasize the tracking of the dominant subspaces of the channel vector autocorrelation matrix. This approach reduces the weight noise in the de-emphasized subspaces, hence enhancing

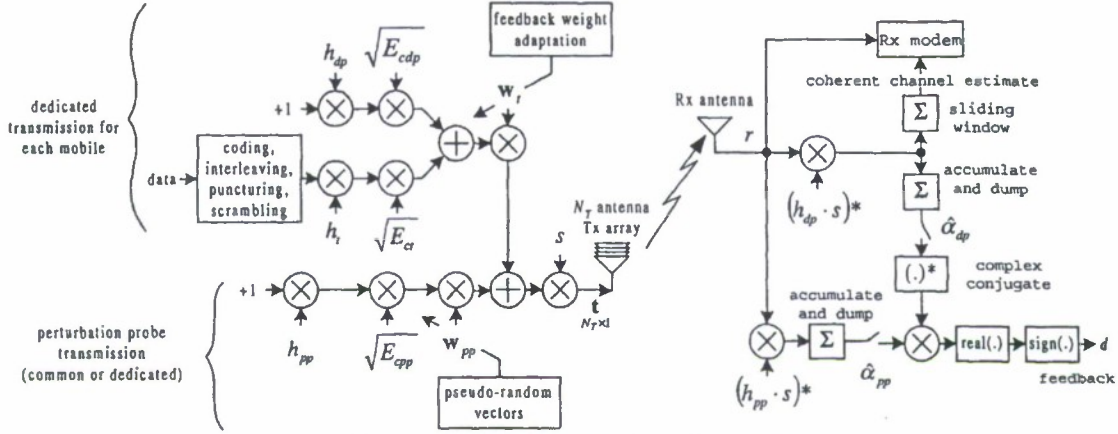


Figure 1: Transmitter and receiver of the system

performance. The performance of the gradient feedback algorithm in the presence of realistic channel estimation, feedback errors and latency is evaluated by simulation of the algorithm in the cdma2000 cellular system.

2. Stochastic Gradient Algorithm Description

2.1. General Perturbation Formulation

A unified perturbation feedback gradient extraction formulation is presented. This general formulation is applicable to systems using dedicated or common perturbation transmission, and incorporates subspace weighting through weighting matrices applied before probing perturbation transmission (Φ_1) or after receipt of the feedback (Φ_2), which can enhance performance as will be described below. The transmitter and receiver block diagrams of the system are shown in Figure 1. For N_T transmit antennas, the transmitted vector sequence \mathbf{t} for a single receiver is given by the sum of a test perturbation probe, coded traffic, and a dedicated pilot, as follows

$$\mathbf{t}_{N_T \times 1}(n) = s(n) \cdot \left(\mathbf{w}_{pp} \left(\left\lfloor \frac{n}{M} \right\rfloor \right) \sqrt{E_{cpp}^{(T)}} h_{pp}(n) + \mathbf{w}_t \left(\left\lfloor \frac{n-D}{M} \right\rfloor \right) \left(\sqrt{E_{ct}^{(T)}} h_t(n) \cdot x \left(\left\lfloor \frac{n}{N} \right\rfloor \right) + \sqrt{E_{cdp}^{(T)}} h_{dp}(n) \right) \right) \quad (1)$$

where n is the CDMA chip rate index. Nyquist filtering and sampling is assumed. The traffic data symbols $x(\cdot)$ consist of N CDMA chips, $s(n)$ is the pseudo-random chipping sequence, and \mathbf{w}_t is the norm constrained traffic transmission weight vector. $E_{ct}^{(T)}$, $E_{cdp}^{(T)}$ and $E_{cpp}^{(T)}$ are the transmission energies per chip, and $h_t(n)$, $h_{dp}(n)$ and $h_{pp}(n)$ the orthogonal codes (e.g. Walsh-Hadamard codes), of the traffic, dedicated pilot and perturbation probe channels respectively. The perturbation under test is the norm

constrained perturbation weight vector \mathbf{w}_{pp} . The pilot sequence is +1 modulated by the pilot orthogonal code and the pseudo-random chipping sequence, which provides a known pattern for channel estimation at the receiver. The probing period is M chips, with a D chip delay in receiving and implementing the mobile station feedback to update \mathbf{w}_t . The perturbation test vector \mathbf{w}_{pp} is generated from $\mathbf{p} \sim \mathcal{N}(0, 2\mathbf{I})$, which is an uncorrelated Gaussian vector, with a Hermitian positive semi-definite pre-weighting matrix Φ_1 as follows

$$\mathbf{w}_{pp}(m) = \frac{\Phi_1 \mathbf{p}(m)}{\|\Phi_1 \mathbf{p}(m)\|} \quad (2)$$

The conjugated complex flat Raleigh fading channel vector gain is $\mathbf{c}(n)$. With zero mean Gaussian noise $\mathbf{v}(n)$ derived from band limited white noise of density N_0 the received signal is

$$\mathbf{r}(n) = \mathbf{c}^H(n) \mathbf{t}(n) + \mathbf{v}(n) \quad (3)$$

The receiver uses the pilot transmission to estimate the channel for demodulation. The feedback is generated indicating whether a positive or negative adaptation of \mathbf{w}_t in the direction of \mathbf{w}_{pp} is preferable, i.e. the received power would be increased by addition of $\pm \mathbf{w}_{pp}$. For each measurement interval, the receiver generates pilot channel and perturbation channel estimates

$$\hat{\alpha}_{dp}(m) \equiv \alpha_{dp}(m) = \sqrt{E_{cdp}^{(r)}} \cdot \mathbf{c}^H \left(mM + \frac{M}{2} \right) \mathbf{w}_t(m) \quad (4)$$

$$\hat{\alpha}_{pp}(m) \equiv \alpha_{pp}(m) = \sqrt{E_{cpp}^{(r)}} \cdot \mathbf{c}^H \left(mM + \frac{M}{2} \right) \frac{\Phi_1 \mathbf{p}(m)}{\|\Phi_1 \mathbf{p}(m)\|} \quad (5)$$

The feedback decision bit, ± 1 , is generated according to

$$d = \text{sign} \left(\lim_{\Delta \rightarrow 0} \left(|\hat{\alpha}_{dp} + \Delta \hat{\alpha}_{pp}|^2 - |\hat{\alpha}_{dp} - \Delta \hat{\alpha}_{pp}|^2 \right) \right) = \text{sign}(\text{Re}(\hat{\alpha}_{dp} \cdot \hat{\alpha}_{pp}^*)) \quad (6)$$

where the $\text{sign}(\text{Re}())$ operation would be performed on a summation over the channel estimates for multiple Rake fingers in the case of resolvable multipath. The update at the base station upon receipt of this feedback bit $d(m)$ is given by an update with the preferred sign of w_{pp} weighted by a Hermitian positive semi-definite post-weighting matrix Φ_2

$$w_i(m+1) = \frac{w_i(m) + d(m)\beta \cdot \Phi_2 \frac{\Phi_1 p(m)}{\|\Phi_1 p(m)\|}}{\|w_i(m) + d(m)\beta \cdot \Phi_2 \frac{\Phi_1 p(m)}{\|\Phi_1 p(m)\|}\|} \quad (7)$$

where β is an algorithm parameter which controls the rate of the adaptation, with a larger value giving a faster but noisier adaptation. The trace of Φ_2 is constrained to be N_T so that β captures the magnitude of the update. The mobile's perturbation decision is based on the observed differential change in pilot power, so that the update is characterized by the weighted gradient of the received power with respect to the weight vector.

2.2. Gradient Extraction

The performance metric of the system is the ratio of the useful power delivered to the mobile receiver to the power transmitted, given the weight vector. For a conjugated flat fading channel gain vector c this is given by the inverse cost parameter J

$$J \equiv w_i^H c c^H w_i \quad (8)$$

This is maximized by the "matched filter" weights for arbitrary phase rotation ϕ

$$\arg\left(\max_{\|w_i\|=1} (w_i^H c c^H w_i)\right) = \frac{c}{\|c\|} \cdot e^{j\phi} \quad (9)$$

The non-weighted gradient extracted in the receiver is

$$g \equiv \nabla_w(J) = 2c^H w_i c \quad (10)$$

It can be shown that the update of (7) has an expected value given by the normalized gradient vector. In particular, excluding the update normalization applied in (2) the update is

$$z \equiv d(m) \cdot \Phi_2 \Phi_1 p(m) \quad (11)$$

If receiver estimation and feedback errors are ignored this non-normalized update vector z has 1st and 2nd moments

$$E(z) = \sqrt{\frac{2}{\pi}} \cdot \frac{\Phi_2 \Phi_1^2 g}{\|\Phi_1 g\|} \quad (12)$$

$$\begin{aligned} E((z - E(z))(z - E(z))^H) \\ = 2\Phi_2 \Phi_1^2 \Phi_2 - \frac{2}{\pi} \frac{\Phi_2 \Phi_1^2 g g^H \Phi_1^2 \Phi_2}{g^H \Phi_1^2 g} \end{aligned} \quad (13)$$

The normalization applied from (2) in (7) is appropriate

because the feedback from the receiver does not indicate the magnitude of the gradient, but merely the direction. Closed form moments of the normalized perturbation update are not available, but they will clearly have characteristics similar to (12) and (13).

2.3. System Performance Considerations

From (10) and (12) we note that the extracted gradient estimate can include a component radial with the current weight vector w_i . Adaptation in this direction would increase the weight norm and hence transmission power, but such is prevented in the weight normalization of the update in (7). Hence, this component of the adaptation is not useful. The receiver cannot distinguish whether a perturbation vector would hypothetically modify the transmission power, so it is useful for the transmission system to avoid the use of perturbation probe vectors which have this effect. This is accomplished through the use of orthogonal projection in the pre-weighting matrix

$$\Phi_1 = I - \frac{w_i w_i^H}{w_i^H w_i} \quad (14)$$

This technique requires a dedicated transmission, since the pre-weighting is a function of the mobile-specific weight vector. As has been discussed, in a CDMA system there may be many signals transmitted for many mobile receivers, each requiring the transmission of encoded data, pilot, and perturbation. In the context of many users, it can be beneficial to sacrifice the performance enhancement from (14) in order to use a single common perturbation transmission. This tradeoff will be considered in detail below. Then the traffic and pilot transmissions of (1) would be dedicated to the specific mobile while the perturbation of (1) would be common to all mobiles. In this case the pre-weighting Φ_1 must be equi-diagonal since it cannot include mobile specific information, so that $\Phi_1 = I$.

2.4. Enhancing Performance in Correlated Fading

Consider a Raleigh fading channel with correlation across the transmission antennas given by the autocorrelation matrix R .

$$R \equiv E(c(n)c^H(n)) \quad (15)$$

If this correlation is not equi-diagonal the adaptation algorithm can take advantage of knowledge of the correlation. Here it is assumed that a common perturbation is preferable, so that a mobile specific Φ_1 cannot be used. In other circumstances, correlation pre-weighting of a dedicated perturbation may be appropriate. The correlation is used by selecting the update weighting to emphasize those subspaces most occupied by c .

$$\Phi_2 = \frac{N_T \cdot \left(\frac{N_T \cdot \hat{\mathbf{R}}}{\text{tr}(\hat{\mathbf{R}})} + b\mathbf{I} \right)^p}{\text{tr} \left(\left(\frac{N_T \cdot \hat{\mathbf{R}}}{\text{tr}(\hat{\mathbf{R}})} + b\mathbf{I} \right)^p \right)} = \frac{N_T \cdot \mathbf{Q} \left(\frac{N_T \cdot \Lambda}{\text{tr}(\Lambda)} + b\mathbf{I} \right)^p \mathbf{Q}^H}{\text{tr} \left(\left(\frac{N_T \cdot \Lambda}{\text{tr}(\Lambda)} + b\mathbf{I} \right)^p \right)} \quad (16)$$

where $\mathbf{Q}\mathbf{\Lambda}\mathbf{Q}^H$ is the eigendecomposition of $\hat{\mathbf{R}}$, which is an estimate at the transmitter of \mathbf{R} . The parameters p and b define the degree of subspace emphasis and ensure global tracking ability respectively. In the limit as $p \rightarrow \infty$, the gradient algorithm is not adapting to the fading but rather transmitting a fixed beam pattern into the principle subspace of \mathbf{R} , which is the optimal static beam pattern. In the limit as $p \rightarrow 0$ no emphasis is applied, $\Phi_2 = \mathbf{I}$, and the adaptation equally tracks all subspaces of \mathbf{R} . Setting $p = 1/2$ provides an update autocorrelation the same as the channel autocorrelation if $b = 0$. In the limit as $b \rightarrow 0$ null spaces can never be tracked or recovered if the channel statistics change, and a non-zero b allows for some minimal tracking of weak modes. Matrix trace normalization is applied throughout as the least complex normalization for practical implementation, which ensures that the update weight β captures the update magnitude and that the meaning of the identity weighting b is well defined.

The autocorrelation estimate used in (16) can be generated in several ways:

- estimate the channel fading correlation on the uplink and apply directly, or with some modification for the duplex frequency offset, as the downlink estimate.
- parameter estimation, estimating the mean angle of arrival and angular spread on the uplink, and applying these on the downlink through a parameterized correlation matrix calculation.
- weighted averaging of the outer product of the applied forward link weight vectors.
- any linear or other combination of a-c.

The method evaluated in this study is the exponentially weighted summation of the applied forward link weight vectors, with a forgetting factor a . Hence,

$$\hat{\mathbf{R}}(0) = \frac{1}{N_T} \mathbf{I} \quad (17)$$

$$\hat{\mathbf{R}}(m) = a \cdot \hat{\mathbf{R}}(m-1) + (1-a) \cdot \mathbf{w}_i(m) \mathbf{w}_i^H(m) \quad (18)$$

The eigendecomposition of (16) is a convenient way to allow arbitrary power of p , but for integer p this clearly does not need to be explicitly performed. Note that the diagonal weighting by b in (18) not only ensures some minimal adaptation in all subspaces, but also ensures good conditioning for eigendecomposition.

3. System Evaluation

3.1. System Description

Several configurations of the system utilizing 4 transmit antennas are evaluated in order to explore the efficacy of the algorithm and its permutations. The simulated system is a cdma2000 system with a 9600bps data rate, using the constraint length 9, rate 1/4 tailed convolutional code (with octal generator functions: $g_0=765$, $g_1=671$, $g_2=513$, $g_3=473$). The frame length is 20ms (192 bits, including the 8 bit coding tail). The system chip rate is 1.2288Mcps, with no symbol puncturing for reverse link power control. A single fading path is implemented with no inter-chip interference, so that perfect code multiplexed channel orthogonality is maintained. The dedicated pilot power is 5dB lower than the traffic power, $E_{cdp} = E_{cr} - 5\text{dB}$.

Standard cdma2000 forward power control is implemented with a 800bps inner feedback loop controlling $\pm 0.5\text{dB}$ steps at the base station and with an outer loop adjusting the mobile station target E_b/N_0 to provide the desired frame error rate. Note that the dedicated pilot and dedicated perturbation transmissions are power controlled with the traffic in order to maintain their relative powers; common perturbation is not power controlled.

The antenna adaptation feedback rate is 1600bps. The performance is evaluated for both dedicated and common perturbation transmission. The perturbation vector is transmitted with energy per chip E_{cpp} at the same level as the dedicated pilot for dedicated perturbation tests and at a fixed level for common perturbation tests. The channel correlation weighting was performed using pre- or post-feedback weighting with Φ_1 or Φ_2 using a diagonal weighting $b=0.2$ and exponentiation $p=0.5$, which was updated every frame (20ms) while the exponentially weighted estimation of (18) was updated every feedback interval (625 μs). The correlation estimation used a 1s time constant, so that $a=1599/1600$.

There are three separate channel estimation processes at the receiver, for demodulation, power control feedback and antenna feedback. All channel estimates are generated by rectangular weighted correlations of the pilot sequence. For coherent demodulation, the channel estimation is limited by the need to track rapid fading channels with a Doppler frequency up to 100Hz and the correlation is performed as a sliding window over 4096 chips (3.33ms). For antenna feedback the estimate is an accumulate-and-dump correlation of the dedicated pilot and perturbation sequences over the 768 chip (625 μs) measurement period, and the decision is made according to (6). Finally, the power control feedback is based on estimating E_b/N_0 from a pilot channel accumulate-and-dump correlation over the 1536 chip (1.25ms) power control interval.

3.2. Channel Model

The channel model is frequency flat Rayleigh fading with a Jakes time-spectrum, with evaluation of both correlated and uncorrelated antenna fading. For the correlated fading channel, the 4 antennas are configured as a uniform array over 5 wavelengths, so that the spacing is 1.67λ . The angle of arrival "power azimuth spectrum" at the base station for the correlated channel is distributed according to the Laplacian distribution [11][12]

$$P(\theta) = \frac{1}{\sigma_\theta \sqrt{2}} \exp \left[\frac{-\sqrt{2}|\theta - \bar{\theta}|}{\sigma_\theta} \right] \quad (19)$$

where the mean angle of arrival $\bar{\theta}$ is 0° (broadside) and the RMS angular spread σ_θ is 5° , representing an outdoor environment with elevated base station antennas [12]. With this configuration, the fading correlation at the antennas is

$$\mathbf{R}_{4 \times 4} = \text{toeplitz}([1 \quad 0.7066 \quad 0.3747 \quad 0.2105]) \quad (20)$$

where $\text{toeplitz}()$ denotes a hermitian-symmetric Toeplitz extension from the first row. The eigenvalues of \mathbf{R} are

$$\Lambda = \text{diag}([2.5680 \quad 0.9558 \quad 0.3492 \quad 0.1271]) \quad (21)$$

which indicates that a 4.09dB performance enhancement over single antenna transmission is obtained with the optimal static weights, which are given by the principle eigenvector and would provide aperture gain without diversity gain. Note that this is less than the 6.02dB aperture gain which might be expected from 4 antennas.

3.3. Simulation

Simulations were performed for both a non-degraded condition and a degraded condition. The non-degraded condition provided no feedback bit errors, zero feedback latency and perfect channel estimation. The degraded condition provided noisy channel estimation, as described above, with 5% feedback bit error rate and a 312.5μs feedback latency. The legend for all of the simulation plots is shown in Figure 2. The transmission systems are labeled A-F as follows:

- A: $N_T=1$, single antenna transmission
- B: $N_T=4$, perfect transmission weights, $\mathbf{w}_i \sim \mathbf{c}$
- C: $N_T=4$, gradient feedback, $\Phi_1=\mathbf{I}$, $\Phi_2=\mathbf{I}$
- D: $N_T=4$, gradient feedback, $\Phi_1=\mathbf{I}$, $\Phi_2 \sim (\hat{\mathbf{R}} + 0.2\mathbf{I})^{1/2}$
- E: $N_T=4$, gradient feedback, $\Phi_1=\mathbf{I}-\mathbf{w}_i\mathbf{w}_i^H$, $\Phi_2=\mathbf{I}$
- F: $N_T=4$, gradient feedback, $\Phi_1=(\mathbf{I}-\mathbf{w}_i\mathbf{w}_i^H) \cdot (\hat{\mathbf{R}} + 0.2\mathbf{I})^{1/2}$, $\Phi_2=\mathbf{I}$

The channel estimation conditions evaluated are

- (i) perfect channel estimation, no feedback errors, zero feedback latency, E_{cnp} excluded from dedicated power
- (ii) common perturbation channel ($E_{cnp}^{(T)} \cdot |c|_2^2/N_o = -7\text{dB}$), realistic channel estimation and feedback degradation

- (iii) dedicated perturbation channel ($E_{cnp}=E_{ct} - 5\text{dB}$), realistic channel estimation and feedback degradation

In practice, E and F require dedicated perturbation transmission, so that E(ii) and F(ii) are not possible. The expected power gain of an antenna-to-antenna path is portrayed as $|c|_2^2 = E(|c|_2^2)$ for arbitrary channel vector element i , which allows fair comparison of required transmission power.

The simulation result figures show the mean channel normalized inverse cost metric or "mismatch loss", $J' = J/|c|_2^2$, and the mean transmit power to attain 1% FER (frame error rate) with closed loop power control. The transmit energy per bit is $E_{bd}^{(T)}$, which is the dedicated energy and includes the traffic channel, the dedicated pilot and the perturbation channel (if dedicated). Hence, the common perturbation test cases (i)(ii) include a 1.19dB penalty from the pilot, while the dedicated perturbation test cases (iii) include a 2.13dB penalty from the pilot and perturbation. In this way, the overall transmission cost of the system operation is explicit. As an exception for (i), the single transmit antenna case includes no pilot penalty, since a single common pilot could be used for all mobiles.

3.4. Discussion

A sweep of β is first performed with perfect channel estimation for the different pre/post weighting approaches in uncorrelated (Figure 2, Figure 3) and correlated fading (Figure 4, Figure 5). These are all performed at a moderately low fading rate of $f_d=10\text{Hz}$, and show an optimal value for β of 0.25, which is then used for all subsequent results. In these plots the perturbation is considered common (power excluded from E_{bd}) and we see that with perfect channel estimation the pre-weighting techniques with Φ_1 outperform post-weighting with Φ_2 , as we should expect because the weighting is contained in the signal measured by the receiver, rather than distorting the update after the receiver measurement. The application of correlation weighting does not have much effect on the performance at this Doppler frequency, though a performance enhancement is visible in J' in the correlated fading case (Figure 5). Note that the explicit inclusion of the dedicated pilot power in the multiple antenna cases makes it possible for the multiple antenna algorithm to perform worse than the single antenna transmission (Figure 2). In the uncorrelated case we see that as $\beta \rightarrow 0$, so that no tracking takes place, the mean mismatch loss J' goes to -6.02dB (Figure 3), as is to be expected because the potential 4x array power gain is lost. In the correlated case, as $\beta \rightarrow 0$ the mean power delivery is higher, as the antenna weights tend to stabilize in a desirable subspace (Figure 5).

In Figure 6 and Figure 7 the impact of channel estimation errors, feedback bit errors, feedback latency and dedicated versus common perturbation transmission

are considered for the straightforward non-weighted adaptation approach with $\Phi_1 = \Phi_2 = I$. For common perturbation we see that the degradations cause a 1 to 2dB loss in performance from the non-degraded gradient feedback result, while the use of a dedicated perturbation causes an additional 1 to 2dB of loss, which includes effects of both the additional perturbation power and the reduced reliability of the feedback decision estimate. Considering the ~2dB gain from the dedicated techniques using Φ_1 in Figure 2, and Figure 4, it is not clear whether the dedicated or common technique will provide the best overall performance. This is considered in Figure 8 and Figure 10, where results show the relative performance of each algorithm, explicitly including the cost of the dedicated perturbation where such is required. This shows that the common perturbation techniques outperform the equivalent dedicated perturbation techniques by ~1dB. For the pre-weighting techniques (E, F) the loss of performance in going from the non-degraded case (perturbation power excluded) to the degraded case (perturbation power included) is large, about 4dB. This far exceeds the loss for the common perturbation techniques (C, D) of about 1.5dB, and the difference exceeds the 0.94dB excess power penalty from the inclusion of the perturbation power. This indicates that a substantial portion of the loss in the dedicated perturbation approach is incurred from estimation errors in the generating the feedback decision. This is confirmed by the substantial performance loss of the dedicated techniques with degradations in terms of J' in Figure 9 and Figure 11.

The subspace weighting techniques using the estimate of the channel autocorrelation function are seen to improve the tracking performance in the correlated channel condition. In the uncorrelated case the estimation error in the correlation matrix distorts the adaptation but causes only a minimal degradation (Figure 8). In the correlated case the performance is enhanced by up to ~1.5dB at the higher fading frequencies (Figure 10). In the case of rapid fading where the gradient feedback algorithm is not capable of tracking the dynamic channel, the correlation weighting approach allows the antenna to transmit into the principle subspaces of the correlation matrix. In this way, the transmission can be steered in the general direction of the receiver without complete tracking of the independent fades. Diversity is lost when the adaptation cannot track the channel, but a beam steering gain is maintained.

4. Conclusion

A generalized perturbation gradient extraction algorithm for adapting transmit antenna weights using feedback has been presented. A subspace weighting approach has been introduced, which has been shown to enhance the tracking performance when the fading at the antennas is correlated.

The feedback gradient algorithm for adapting transmit weights has been simulated in the cdma2000 digital cellular system with realistic channel estimation and degradations and shown to provide significant performance enhancement relative to a single transmit antenna. It was shown that in a CDMA system the technique can be applied with a common perturbation, used by all receivers, and that this will increase overall capacity by reducing the transmission power dedicated to each mobile.

5. References

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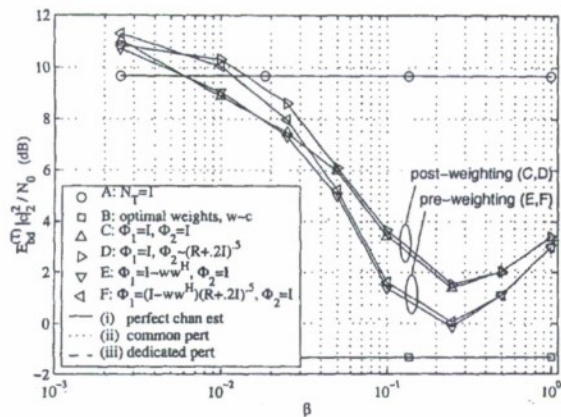


Figure 2: Mean E_d/N_0 vs. β , power controlled to 1% FER. Uncorrelated channel, $f_d=10\text{Hz}$, no degradations

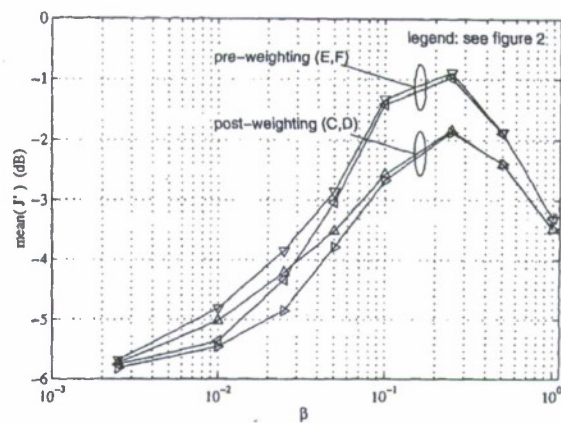


Figure 3: Mean metric J' vs. β . Uncorrelated channel, $f_d=10\text{Hz}$, no degradations.

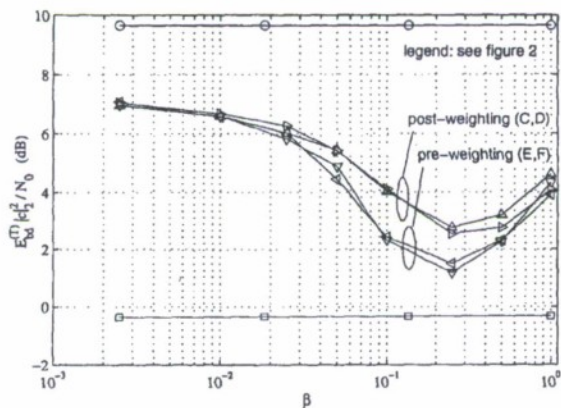


Figure 4: Mean E_d/N_0 vs. β , power controlled to 1% FER. Correlated channel, $f_d=10\text{Hz}$, no degradations

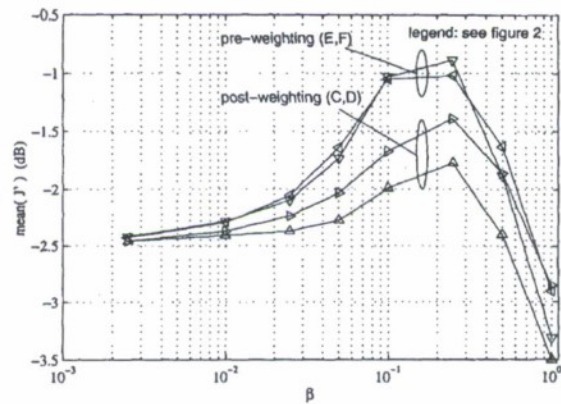


Figure 5: Mean metric J' vs. β . Correlated channel, $f_d=10\text{Hz}$, no degradations.

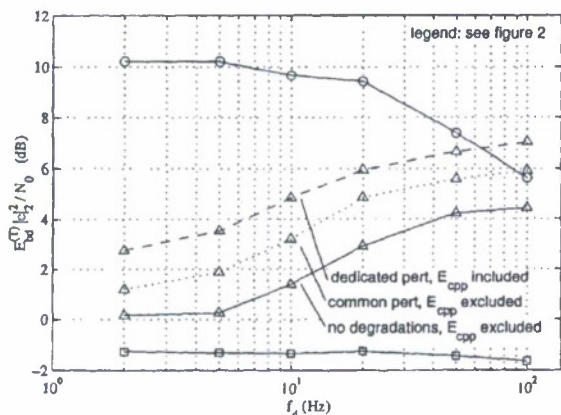


Figure 6: Mean E_d/N_0 vs. f_d , power controlled to 1% FER. Uncorrelated channel, $\beta=0.25$. Compare perfect channel estimation and realistic estimation (common/dedicated pert)

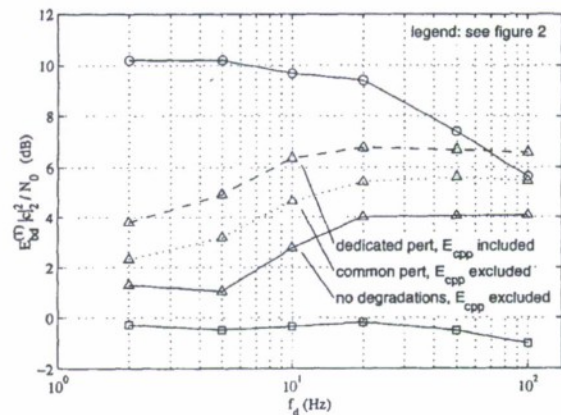


Figure 7: Mean E_d/N_0 vs. f_d , power controlled to 1% FER. Correlated channel, $\beta=0.25$. Compare perfect channel estimation and realistic estimation (common/dedicated pert)

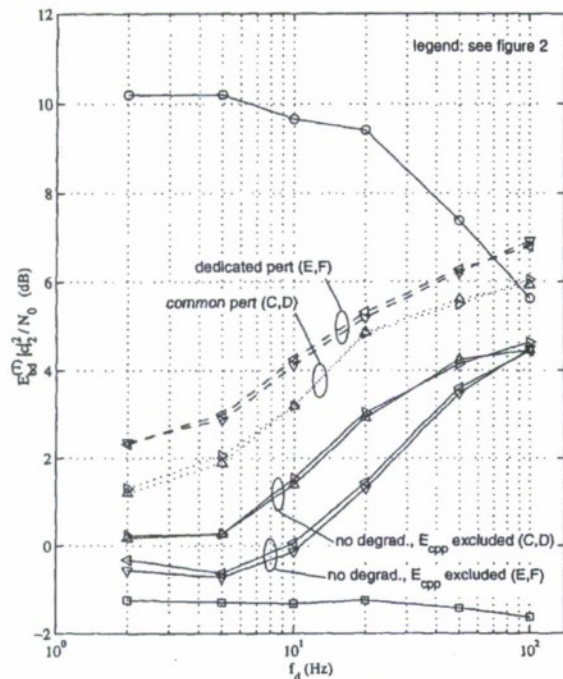


Figure 8: Mean E_d/N_0 vs. f_d power controlled to 1% FER. Uncorrelated channel, $\beta=0.25$, with/without degradations.

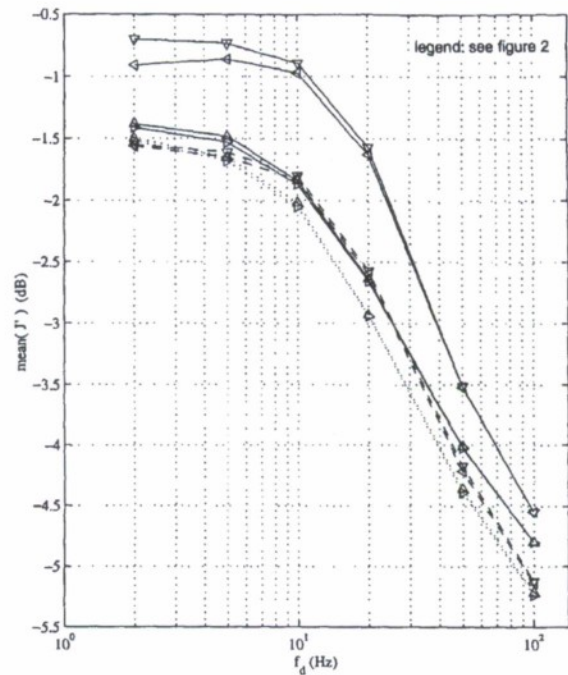


Figure 9: Mean metric J' vs. f_d . Uncorrelated channel, $\beta=0.25$, with/without degradations

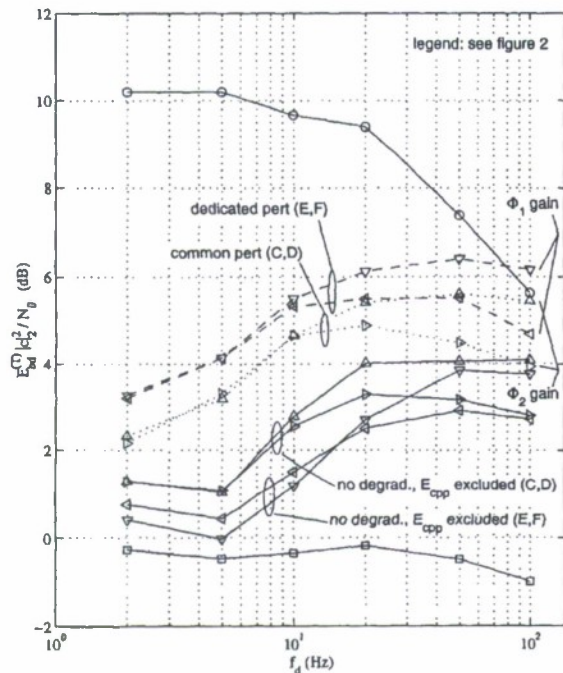


Figure 10: Mean E_d/N_0 vs. f_d , power controlled to 1% FER. Correlated channel, $\beta=0.25$, with/without degradations.

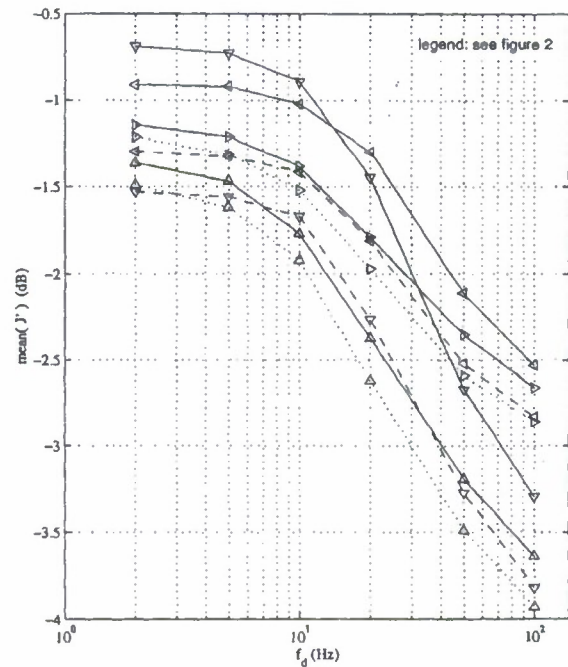


Figure 11: Mean metric J' vs. f_d . Correlated channel, $\beta=0.25$, with/without degradations